

MAXIM Engineering Journal

Volume Twenty-Nine

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News Briefs

■ MAXIM REPORTS RECORD REVENUES AND EARNINGS FOR THE FIRST QUARTER OF FISCAL 1998

Maxim Integrated Products, Inc., reported record net revenues of \$125 million for the first quarter of fiscal 1998 ending September 27, 1997, compared to \$101 million for the same quarter in fiscal 1997. Net income increased to \$40 million compared to \$31.4 million for the first quarter of fiscal 1997. Income per share was \$0.53 per share for the quarter compared to \$0.45 per share for the same period a year ago.

During the quarter, Maxim's cash and short-term investments increased \$51.4 million after purchasing \$10.6 million of common stock and \$10.5 million of capital equipment. Total cash and short-term investments at the end of Q198 equaled \$275.3 million. Annualized return on average stockholders' equity for the quarter was 32%, one of the highest in the industry today.

During Q198, backlog shippable within the next 12 months increased to \$182 million from the \$152 million reported at the end of Q497. Approximately 80% of the Q198 backlog consists of orders that were requested for shipment in Q298 or earlier.

Turns orders received in Q198 were \$50.7 million. (Turns orders are customer orders that are for delivery within the same quarter and may result in revenue within the quarter if the Company has available inventory that matches those orders.)

Worldwide net bookings were higher in Q198 than in Q497, with net bookings in the Pacific Rim and Europe showing the greatest increase. Net bookings for all product areas continue to be strong, particularly for those products based on the Company's high-frequency bipolar technology. In addition, net bookings for product lines focused on our broadest markets (instrumentation and process control) continue to be strong.

Gross margins for the first quarter increased slightly to 66.8% compared to 66.4% in Q497. Increases in production volume and manufacturing productivity continued in Q198. Research and development expense increased to \$15.5 million, 12.4% of net revenues, due to the Company's continued investment in new product development.

In July, *Forbes* Magazine listed Maxim as one of the top ten new issues of the decade. Maxim ranked number seven on this elite list. In September, *Fortune* Magazine featured Maxim as one of the top 100 fastest growing companies in the United States. Maxim was ranked 48th out of the top 100, based on earnings per share annual growth rate. Maxim ranked 8th out of the *Fortune* top 100 companies in terms of net income for a four-quarter period and had the 16th best one-year share price performance of the group.

Jack Gifford, Chairman, President and Chief Executive Officer, commented on the quarter: "During Q198, Maxim continued to see very broad demand for its products. Demand from the Pacific Rim and Europe was particularly strong. Although some parts of Asia are experiencing recessionary problems, Asian customers are demanding Maxim products at a high rate. We attribute this demand to the fact that Maxim ICs are used mainly in such Asian exports as computers and electronic instruments and are not intended for internal consumption. Maxim sales continue to be well balanced with 14% derived from customers located in the Pacific Rim, 17% from Japan, and 25% from Europe. We believe that the consumption of Maxim products by its customers is at record levels worldwide."

Gifford continued: "We are pleased to see the continued recognition of Maxim by the financial community, including recent articles in *Forbes* and *Fortune*. I was particularly proud to see Maxim ranked one of the top ten new IPOs of the decade, considering the stiff competition. This coming year will mark Maxim's first decade as a publicly traded company. During our second decade, we will continue to work hard to outperform our competition through strong management, engineering ingenuity, and high productivity in hopes of continuing recognition by stockholders."

Silicon-bipolar IC facilitates VCO design

The frequency of a voltage-controlled oscillator (VCO) varies with the voltage applied to its tuning port. Operating in a phase-locked loop (PLL), the VCO provides a stable local oscillator (LO) for frequency conversion in superheterodyne receivers. VCOs are also used in transmit chains, where they upconvert the baseband signal to a radio frequency (RF) suitable for transmission over the airwaves (Figure 1).

Design considerations

The VCO designer must consider several important performance parameters:

- Output level in dBm (dB relative to 1mW)
- Output harmonic level in dBc (dB relative to carrier power)
- Tuning sensitivity in Hz/V
- Load pulling of oscillation frequency in Hz p-p (for a given load voltage standing-wave ratio (VSWR) rotated through 360°)
- Frequency pushing, in Hz/V, of bias-supply change
- VCO phase noise, in dBc/Hz, at a given offset frequency

The following paragraphs discuss each parameter in turn.

Output level

In typical superheterodyne receivers, the VCO output must drive a mixer as well as a PLL synthesizer's RF prescaler. This requirement is commonly met with a buffer amplifier, which provides load isolation as well as greater drive capability.

Output harmonic level

Output harmonic level is a measure of the VCO energy at harmonics of the oscillation frequency. These harmonics, common at levels below -15dBc, are generated by the nonlinear self-limiting of active devices in the oscillator. Oscillators with large amounts of excess gain (greater than the amount necessary to offset all losses at resonance) will limit more severely, thereby generating a greater harmonic content in the output waveform. The designer must balance the need to keep harmonic levels low with the need for enough excess gain to ensure oscillator start-up.

Tuning sensitivity

Tuning sensitivity is a system-level parameter that relates the maximum available tuning voltage to the required tuning-frequency range, in units of Hz/V. It is inversely proportional to the loaded Q, which is the loaded oscillator tank's quality factor. Higher tuning sensitivities require oscillators with lower loaded Qs.

The variation of tuning sensitivity over the tuning-frequency range is another important consideration. If a

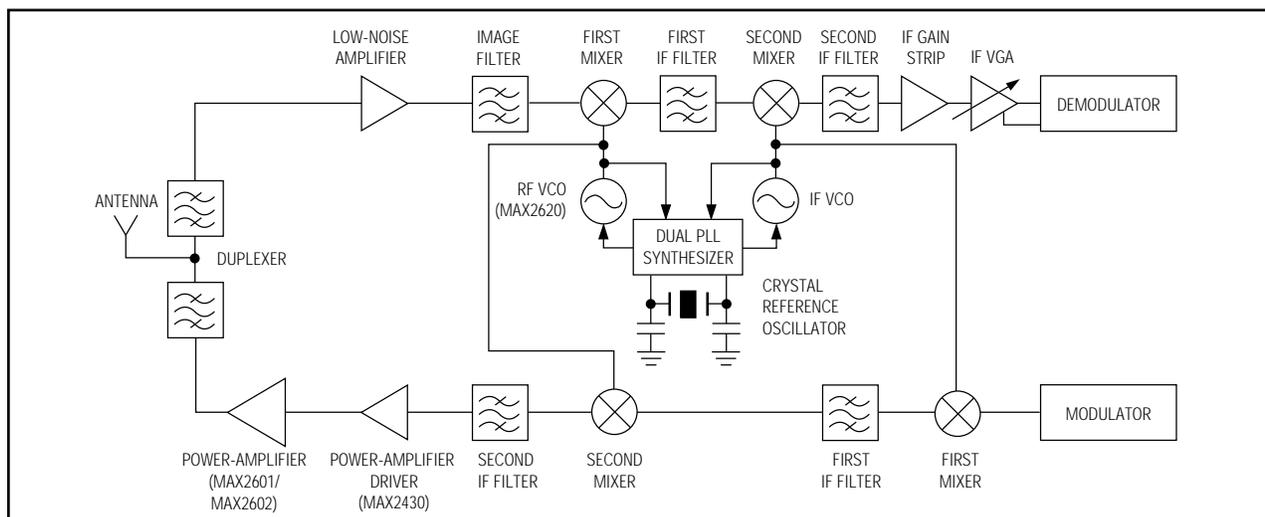


Figure 1. VCOs appear as part of the PLLs in this typical superheterodyne receiver.

VCO's tuning sensitivity varies dramatically over the tuning band, the PLL synthesizer's performance suffers. The VCO is the highest gain device in a typical PLL, with tuning sensitivities in the tens of MHz/V. This amount of gain can cause unwanted modulation sidebands in response to noise at the tuning port; therefore, tuning-port noise must be minimized.

Load pulling

Load pulling measures the sensitivity of a free-running VCO to load variations at the VCO output. Measurement requires a load-impedance mismatch and a variable-length transmission line. Connect the VCO to the mismatched load, and vary the phase angle (between VCO and load) through 360° by changing the length of the transmission line. Measure the resulting peak-to-peak frequency change. VCO load pulling is specified as the maximum peak-to-peak frequency shift at a given load VSWR, rotated through 360°. Equation 1 shows the relationship between load VSWR and load-impedance mismatch:

Equation 1:

$$\text{VSWR} = \frac{1 + |\Gamma_0|}{1 - |\Gamma_0|}, \text{ where } \Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0}$$

where:

VSWR = voltage standing-wave ratio

Γ_0 = load-reflection coefficient: the ratio (at the load) of the incident voltage wave to the reflected wave

Z_L = load impedance

Z_0 = the transmission line's characteristic impedance

Using buffer amplifiers is the most common technique for reducing a free-running VCO's sensitivity to load variations.

Frequency pushing

Frequency pushing measures a free-running VCO's sensitivity to variations in its bias-supply voltage. To measure the VCO's sensitivity, vary the supply voltage over a given range while measuring the VCO frequency. Divide this frequency shift by the voltage change to determine sensitivity in Hz/V. Well-designed VCOs have pushing factors between 5% and 10% of the main tuning-line sensitivity. An example of a device with excellent pushing performance is Maxim's MAX2620 VCO, which has a tuning-port sensitivity of 10.4MHz/V and a pushing sensitivity of only 71kHz/V. Pushing sensitivity for the MAX2620 is less than 1% of the tuning-port sensitivity.

VCO phase noise

Phase noise in a free-running VCO relates the noise-sideband level to the carrier-power level. In a typical measurement, observe the VCO output on a spectrum analyzer while measuring the noise level in a 1Hz bandwidth at a given frequency offset from the carrier. Modern spectrum analyzers equipped with a particular firmware option can generate a graph showing single-sideband phase noise versus offset frequency by taking multiple measurements with various offsets, and making appropriate changes to the internal IF bandwidth in each case.

Oscillators with very low phase noise (crystal oscillators, for example) cannot be measured by a spectrum analyzer because the phase-noise limit for its LO is too high. The 8561 RF spectrum analyzer from Hewlett Packard, for instance, specifies phase-noise limits of -80dBc/Hz at 100Hz, -97dBc/Hz at 1kHz, -113dBc at 10kHz, -113dBc at 30kHz, and -113dBc at 100kHz. A typical crystal oscillator, on the other hand, has 30dB to 40dB less phase noise at each of these offset frequencies. For such high-quality oscillators, an accurate phase-noise measurement requires more sophisticated techniques.

Several key factors affect a free-running VCO's phase noise. All are included in Equation 2, a formula for estimating an oscillator's single-sideband noise.

Equation 2:

$$L(f_M) = 10 \log \left[\frac{1}{2} \left[\left(\frac{f_O}{2Q_L f_M} \right)^2 + 1 \right] \left(\frac{f_C}{f_M} + 1 \right) \frac{FkT}{P_S} \right]$$

where:

$L(f_M)$ = single-sideband phase noise in dBc/Hz, as a function of offset frequency from the carrier

f_O = output frequency in Hz

Q_L = loaded resonator Q (resonator tank circuit with active load and all parasitic elements)

f_C = corner frequency in Hz for flicker noise in the active oscillation device

f_M = offset from the carrier in Hz

P_S = the active oscillation device's oscillation-signal power, in watts

F = the active device's in-circuit noise factor (with the resonator tank and all parasitic elements)

k = Boltzmann's constant: $\sim 1.38 \times 10^{-23} \text{ J/}^\circ\text{K}$

T = temperature in degrees Kelvin ($^\circ\text{K}$)

In this formula, loaded-resonator Q is the dominant design parameter affecting phase noise. Low-noise design dictates that this parameter be maximized to meet tuneability requirements. A high loaded-resonator Q requires the use of resonant-tank components with high unloaded Q . Under these conditions, the tank's load should couple just enough energy to the rest of the circuit to start and sustain oscillations. The resonator's loaded Q can easily be less than a tenth of its unloaded Q .

The corner frequency for flicker noise is device dependent; low-noise design demands devices with a low flicker corner. The flicker-noise corner makes bipolar processes the best choice for low-noise oscillator design. GaAs devices cannot compete because their noise corner is from two to three orders of magnitude greater than that of Si-bipolar devices.

Because the value of the in-circuit noise factor depends on the device as well as its external circuit, low-noise design requires that both be optimized. Adjusting the oscillation-signal power allows some control over phase noise, but the premium on bias current in today's handheld wireless phones usually prohibits large changes in the oscillator section's current drain.

Equation 3 describes phase noise intrinsic to the oscillator. Adding to this are the modulation-noise sidebands produced by noise on the tuning line (see Equation 4).

Equation 3:

$$L_{\text{PUSH}(fm)} = 10 \log \left(\frac{(K_{\text{PUSH}}^2)(VN_{\text{BIAS}}^2(f))}{2f_M^2} \right)$$

Equation 4:

$$L_{\text{MOD}(fm)} = 10 \log \left(\frac{(K_{\text{TUNE}}^2)(VN_{\text{TUNE}}^2(f))}{2f_M^2} \right)$$

where:

$L_{\text{PUSH}(fm)}$ = single-sideband phase noise (in dBc/Hz) due to noise voltage modulating the VCO through the bias line

$L_{\text{MOD}(fm)}$ = single-sideband phase noise (in dBc/Hz) due to noise voltage modulating the VCO through the tuning line

K_{PUSH}^2 = supply-pushing sensitivity, in Hz/V

K_{TUNE}^2 = oscillator tuning gain, in Hz/V

$VN_{\text{BIAS}}^2(f)$ = noise-voltage density on the bias line as a function of frequency (nV/ $\sqrt{\text{Hz}}$)

$VN_{\text{TUNE}}^2(f)$ = noise-voltage density on the tuning line as a function of frequency (nV/ $\sqrt{\text{Hz}}$)

Adding equations 2, 3, and 4 results in equation 5, an estimate of the VCO's total single-sideband phase noise:

Equation 5:

$$L_{\text{TOTAL}(fm)} = 10 \log \left[\frac{1}{2} \left[\left(\frac{f_O}{2Qf_M} \right)^2 + 1 \right] \left(\frac{f_C}{f_M} + 1 \right) \frac{FkT}{P_S} + \frac{(K_{\text{PUSH}}^2)(VN_{\text{BIAS}}^2(f))}{2f_M^2} + \frac{(K_{\text{TUNE}}^2)(VN_{\text{TUNE}}^2(f))}{2f_M^2} \right]$$

Limitations in the previously mentioned VCO parameters can lead to degradation of system-level performance. For example, the power amplifier (PA) in a cellular phone is activated only when a voice signal is present. This switching causes the PA's input impedance to vary considerably, which in turn presents a problem for the RF VCO driving the transmit chain. Unless the VCO is isolated from the load variations (typically by a load buffer), its frequency variations can cause the PLL to slip cycles or even lose phase lock.

Another problem is the drastic change in supply current caused by the PA's off/on cycling. Typical PAs for GSM, DCS1800, and DCS1900 handsets can draw over 1A, and the current switching causes voltage changes on the VCO's bias line. The result of these bias-voltage changes and the pushing factor is unwanted modulation sidebands that fall outside of the PLL synthesizer's loop bandwidth. The VCO's bias voltage must be stabilized to eliminate this problem.

The unfaded bit-error rate (BER) in digitally modulated systems is limited by the net phase noise of all signal generators in the transmit and receive paths, with the RF VCO in the PLL synthesizer (usually) as the dominant contributor. The classic waterfall curve in **Figure 2** shows the effect of phase noise. Beyond a certain level

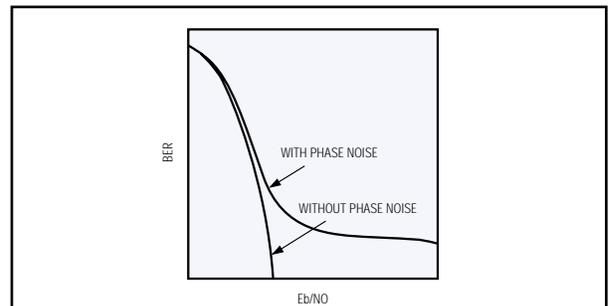


Figure 2. For higher values of energy per bit divided by additive white Gaussian noise density (E_b/N_0), the bit-error rate (BER) is essentially constant.

of E_b/N_0 (E_b is energy per bit; N_0 is additive white Gaussian noise density), the BER remains essentially constant. For a more robust communication link, lower the unfaded BER by reducing phase noise in the PLL synthesizer's RF VCO.

Phase noise is a primary concern for digital-modulation techniques in which information is encoded by modulating the carrier phase. One such technique is quadrature phase-shift keying (QPSK). Analogous to in-phase/quadrature modulation in the analog domain, QPSK allows transmission of a given bit stream at half the data rate by encoding pairs of bits at each of four different phases. Each phase ($\pi/4$, $3\pi/4$, $5\pi/4$, and $7\pi/4$ in **Figure 3a**) is represented as a point in signal space that is spread into a cloud by the presence of additive white Gaussian noise (AWGN) in the system.

Figure 3b shows the same QPSK constellation with the same AWGN, but with 5° of RMS phase variance added. Phase variance deforms the four constellation regions into arcs that reduce the distance between regions. This effect increases the probability of a symbol error at the demodulator, and an increase in symbol errors increases the BER. Thus, the amount of phase variance that can be tolerated depends on the demodulator design and on the performance required in the

communication link. Equation 6 shows the relationship between integrated phase variance and phase noise:

Equation 6:

$$\sigma^2_\phi = \int_{f_1}^{f_2} S_\phi(f) df$$

where:

f_1, f_2 = frequencies over which the integral is evaluated (usually determined by the demodulator design)

σ^2_ϕ = integrated phase variance in radians squared

$S_\phi(f)$ = phase-power spectral density in radians squared/Hz (twice the single-sideband phase noise for small angles)

$\sqrt{\sigma^2_\phi}$ = integrated RMS phase error, in radians

Perhaps the most stringent restraint on an LO's phase noise is imposed by receiver desensitization. This effect occurs in cellular phones and other environments in which the receiver must detect a weak signal in the presence of a strong interferer. In **Figure 4**, a strong nearby interferer mixes with the LO's phase noise to produce noise sidebands that reduce the signal-to-noise ratio at the IF, thus desensitizing the receiver's ability to detect weak signals.

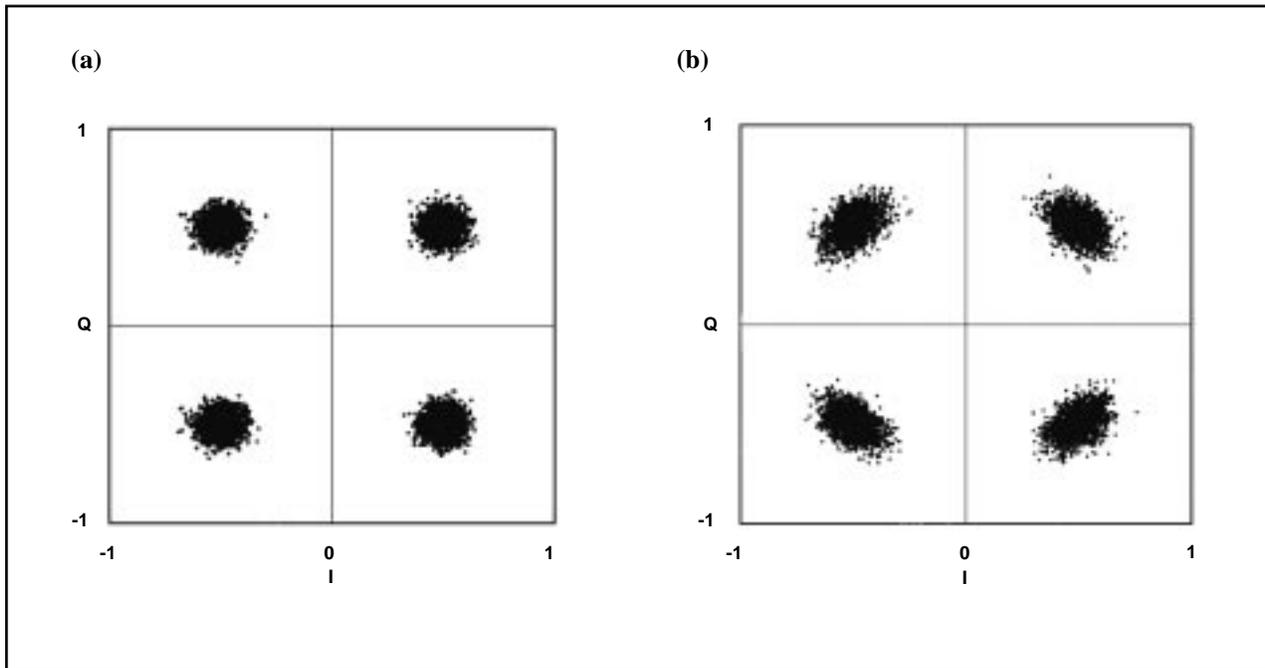


Figure 3. The signal constellation for a QPSK signal with Gaussian noise (a) is degraded by the addition of 5° of RMS phase variance (b), producing a distortion that can raise the BER.

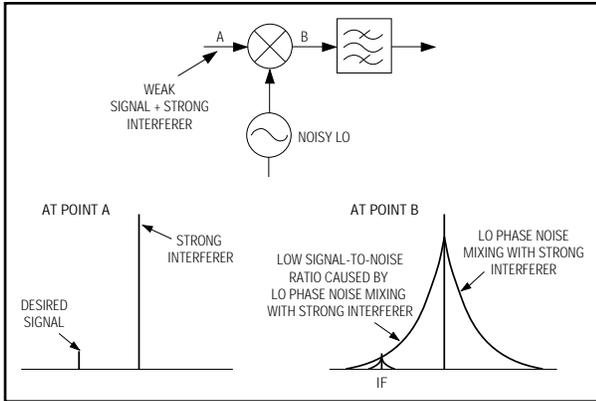


Figure 4. By mixing with the local-oscillator signal, a strong interfering signal generates noise sidebands that mask the signal of interest.

Earlier versions of the low-noise VCO were composed of discrete components: a specialized bipolar transistor with low corner frequency for flicker noise, a bias-voltage supply, and buffer amplifiers to provide load isolation and added output drive. The many passive chip components in the discrete circuit required a lot of PC board space, which is at a premium in today's small wireless handsets.

An integrated solution

Maxim's MAX2620 (Figure 5) integrates all the active functions of a discrete-component approach into a tiny, 8-pin μ MAX package. It includes a critical bipolar transistor with low corner frequency for flicker noise, fabricated in Maxim's exclusive Si-bipolar process featuring a 27GHz f_T . The PC board area saved by higher-level integration simplifies PC board layout and shielding.

Besides the low-noise transistor, the MAX2620 includes a double buffer with two outputs (for load isolation), a bias generator, and convenient shutdown capability. This device operates from a +2.7V to +5.5V single supply and dissipates only 27mW at 3V. When operating at 900MHz, a load VSWR of 1.75:1 rotated over 360° produces a frequency shift of less than 163kHz. The MAX2620's internal bias-voltage generator greatly reduces the effect of bias-voltage variation on the oscillation frequency. At a 900MHz center frequency and a 3V to 4V supply-voltage change, the device achieves a 71kHz/V pushing sensitivity.

The MAX2620 has two outputs. One output, which generates -2dBm into a 50 Ω load, typically drives a mixer's LO input. The other generates -12.5dBm into a 50 Ω load and typically drives an integrated PLL synthesizer's RF prescaler input. Operating at 900MHz with a high-Q external tank circuit, the MAX2620 and its low-noise internal transistor produce low phase noise: -110dBc/Hz at 25kHz and -132dBc/Hz at 300kHz. The external tank allows designers to optimize tuneability and single-sideband phase noise for a given application.

To ensure oscillation start-up, the tank circuit's real-impedance magnitude should equal one-third to one-half of the oscillator device's negative real-impedance magnitude, and the tank's reactive component should be opposite in sign to that of the oscillator device. After start-up, gain compression lowers the oscillator's negative resistance until it achieves equilibrium with that of the resonant tank circuit.

Adding a varactor diode (voltage-tuned variable capacitor) to the tank circuit enables oscillator-frequency

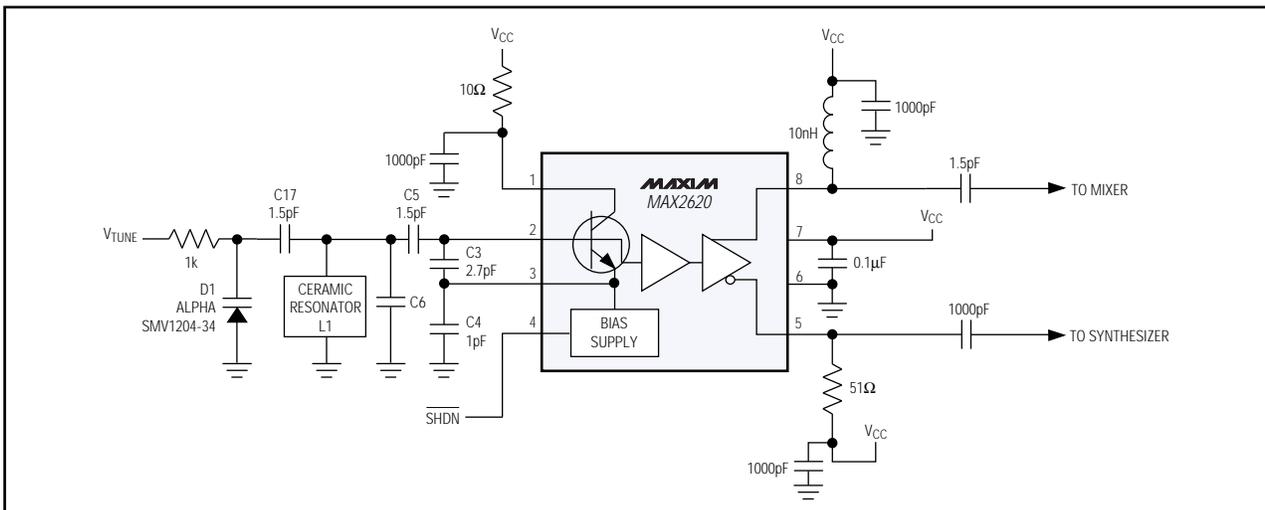


Figure 5. This typical operating circuit shows the use of a MAX2620 in building a VCO.

tuning, as long as the oscillator device exhibits an adequate negative resistance over the desired tuning range. The MAX2620 design is optimized in this respect.

The MAX2620 oscillator is also optimized for operation with low phase noise. Achieving the lowest phase noise possible requires the use of high-Q components such as ceramic transmission-line resonators (typical unloaded Q of 400) and high-Q inductors (typical unloaded Q of 180). To maximize the loaded Q in Figure 5, C5 and C17 should have the lowest value compatible with the desired frequency and tuning range. For 900MHz operation, C6 should be 1pF for the ceramic-resonator circuit and 1.5pF for the inductor circuit. Because a high-Q inductor's unloaded Q is lower than that of a ceramic resonator, the use of high-Q inductors (versus ceramic resonators) tends to degrade phase noise slightly. Phase noise for an inductor-based tank is -107dBc/Hz at 25kHz and -127dBc/Hz at 300kHz.

Both MAX2620 outputs have open collectors that require external components for pull-up to the supply voltage. Resistors of 50Ω match the outputs to a 50Ω system, but resistors rob output power. For maximum output power, use a pull-up inductor as shown at the buffer output in

Figure 5. The inductor circuit's open-collector output impedance should be matched to the desired load impedance through an appropriate matching network.

A key factor in achieving optimum oscillator performance is the PC board layout. To minimize the effect of parasitic elements, remove the PC board ground plane under and around components that make up the resonant circuit. To minimize parasitic inductance, keep trace lengths as short as possible. Connect the decoupling capacitors (pins 1, 4, and 7 to ground) as close as possible to the MAX2620 package, with direct connections to the ground plane. The capacitors in Figure 5 must have an 0805 or smaller footprint.

As a cost-effective, low-power oscillator for the RF VCO in today's wireless headsets, the MAX2620 provides features that once required many discrete parts. Its double-buffered outputs provide load isolation, and its internal regulation cell provides isolation from power-supply fluctuations. Power dissipation with a +3V supply is just 27mW. The MAX2620 achieves very low phase noise, and its external tank lets the designer tailor an oscillator circuit to a given application.

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DESIGN SHOWCASE

SSB modulator draws only 5mA at 2.7V

Single-sideband modulation (SSB) is more efficient than full-amplitude modulation in its use of the frequency spectrum and in its generation of output power. Though not used for data transmission, SSB is still popular for voice transmission at HF and low VHF. The circuit shown in **Figure 1** generates SSB signals from 35MHz to 80MHz by combining wideband, low-voltage op amps with an IC that integrates all the necessary functions. All ICs shown are specified for operation at 3V \pm 10%.

The traditional method for producing SSB is to modulate a carrier, and then filter the output to remove the unwanted sideband and carrier frequencies. This method is sometimes considered wasteful, because it dumps as much as two-thirds of the generated power into a filter. (However, because filtering is not always performed at the output stage, the system doesn't necessarily waste two-thirds of its transmitted power.)

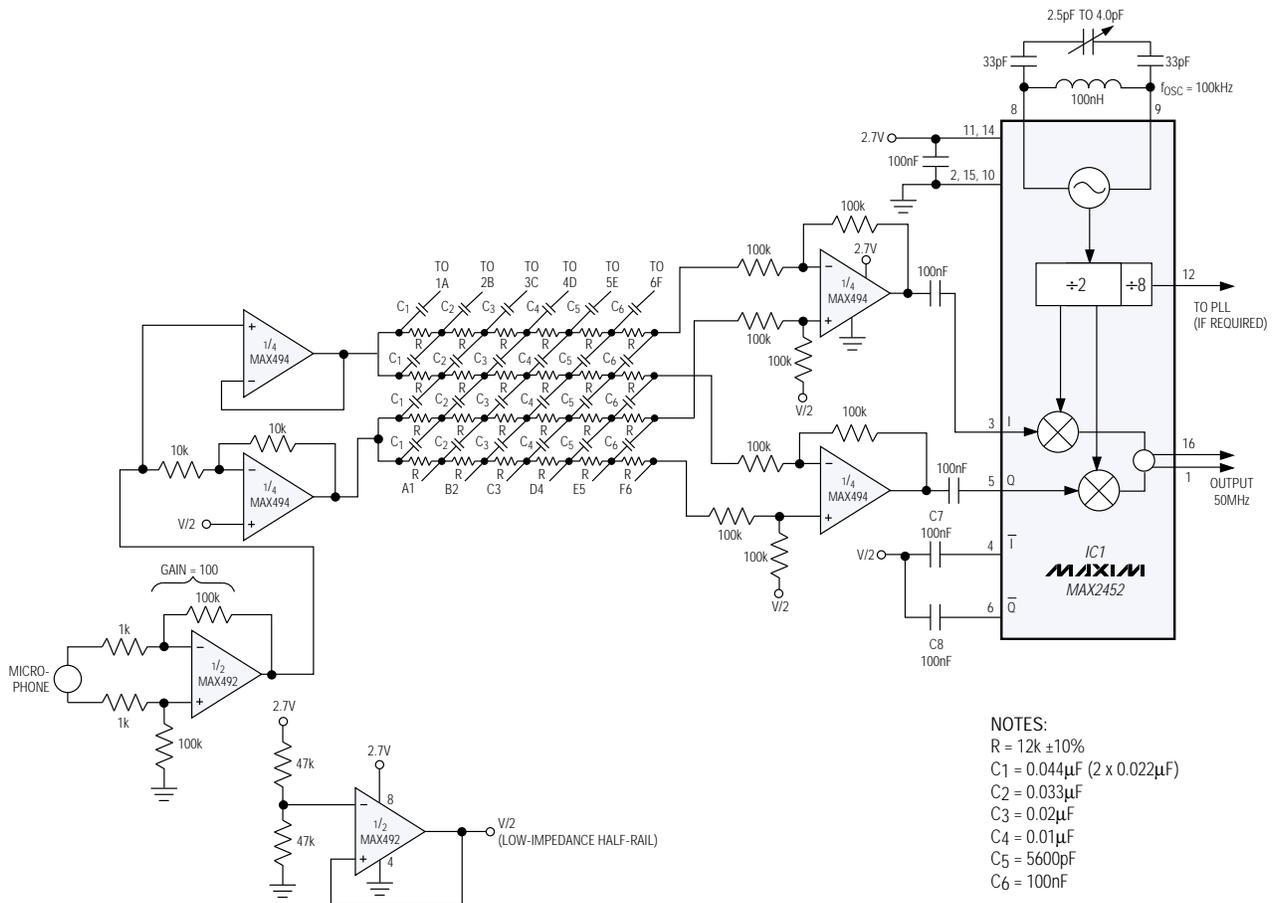


Figure 1. This SSB modulator generates the lower sideband of a high-frequency carrier modulated by an applied audio signal.

An alternative method for generating SSB is to use the phasing (algebraic) method. In this approach, two modulators (mixers) produce the desired sideband while suppressing the unwanted carrier and other sideband. Two modulators for this purpose, normally used for in-phase (I) and quadrature (Q) modulation in a QAM signal, are available in IC1. The resulting circuit offers several advantages:

- Low-power, low-cost operation
- Output signal (35MHz to 80MHz) includes the 4m and 6m amateur radio bands
- User can shift from upper- to lower-sideband operation by reversing two pairs of connections (rather than changing a filter)
- No filter required
- One IC provides the required tank oscillator, two modulators, and a summing amplifier

The circuit requires no filter for carrier and sideband-frequency suppression because frequency cancellation is inherent in the modulation process. Suppose, for example (ignoring signal magnitudes), that the carrier signal is $\sin\omega_C t$ and the modulating signal is $\sin\omega_M t$. Modulation (mixing) means multiplying the carrier and modulating signals, as follows:

$$[\sin\omega_M t] [\sin\omega_C t] = 0.5\cos(\omega_M - \omega_C)t - 0.5\cos(\omega_M + \omega_C)t$$

Adding 90° of phase shift to either quantity produces the cosine: $\sin(\omega_C + 90^\circ)t = \cos\omega_C t$, and $\sin(\omega_M + 90^\circ)t = \cos\omega_M t$. Shifting each of these inputs ($\sin\omega_M t$ and $\sin\omega_C t$) by 90° and then multiplying them in a separate modulator results in the following:

$$[\cos\omega_M t] [\cos\omega_C t] = 0.5\cos(\omega_M - \omega_C)t + 0.5\cos(\omega_M + \omega_C)t$$

Note that the lower sideband, which appears as $\cos(\omega_M - \omega_C)t$ at IC1's output, is the sum of these two modulator outputs. The upper sideband, which appears as $\cos(\omega_M + \omega_C)t$, is the difference between the modulator outputs.

The RC phasing network was chosen for simplicity, rather than low component count. Using 5% components, the network produces a response of 300Hz to 3500Hz with $<1^\circ$ of phase-shift error and $<0.2\text{dB}$ of magnitude error. IC1's suppression of unwanted carrier and sideband frequencies (-35dB) is about 5dB less than expected when using commercial equipment, but is not unreasonable for output power levels below 5W. This suppression performance depends somewhat on the presence of capacitive terminations (C4 and C5) for the unused modulator inputs \bar{I} and \bar{Q} . The output stage (not shown) can be a single-transistor buffer, a class C power amplifier, or whatever the application requires.

For simplicity, the circuit is shown operating with IC1's internal free-running oscillator. This arrangement is insufficiently stable; to compensate for this instability, either provide an external source, or connect the oscillator as part of an external phase-locked loop, as explained in the data sheet for the IC (in this case, the MAX2452). An external source can greatly extend the transmit-frequency range.

The circuit was measured while operating with a 142MHz oscillator frequency and a 71MHz carrier. Its -27dB carrier suppression is 8dB short of the typical suppression specified in the MAX2452 data sheet, but is acceptable for a circuit that drives IC1's I and Q inputs in single-ended mode. (Driving them differentially improves performance.) Sideband suppression was at least -36dB (the test setup's noise floor).

A related idea appeared in the 6/5/97 issue of EDN.

DESIGN SHOWCASE

Adjustment-free inclinometer operates on +2.7V

Figure 1 is an inclinometer (tilt-measuring circuit) whose sensor (SN1) is filled with liquid electrolyte. Acting as a potentiometer, the inclinometer produces a voltage proportional to tilt on its center electrode. Because the liquid is subject to electrolysis, the sensor's forcing voltage must be AC with an average DC component of zero. IC1 is an 8-channel, 12-bit analog-to-digital converter (ADC) that digitizes the sensor output for use by IC2, the microcontroller (μC).

Conditioning circuitry for this sensor type usually includes op amps, analog switches, and potentiometers. Because potentiometer settings drift with time and temperature, such systems require periodic recalibrations based on a precise and tedious procedure. The synchronous approach shown in Figure 1 not only

eliminates the need for calibration, but it also operates from a single-supply voltage as low as +2.7V.

Two CMOS port pins on the μC generate 50Hz square waves, 180° out-of-phase, as an AC drive for the sensor. When the sensor is level, its center-electrode voltage (filtered by R3/C4 and fed to the ADC) is midway between these drive-electrode voltages, which are approximately V_{CC} and 0V. Each port pin has a finite resistance and resultant voltage drop. To compensate for the resulting inaccuracies, voltage divider R4/R5 samples the drive signal's mid-level voltage and feeds it to channel 2 on the ADC. This voltage remains constant, but the center-electrode signal varies above or below mid-level according to the direction of tilt.

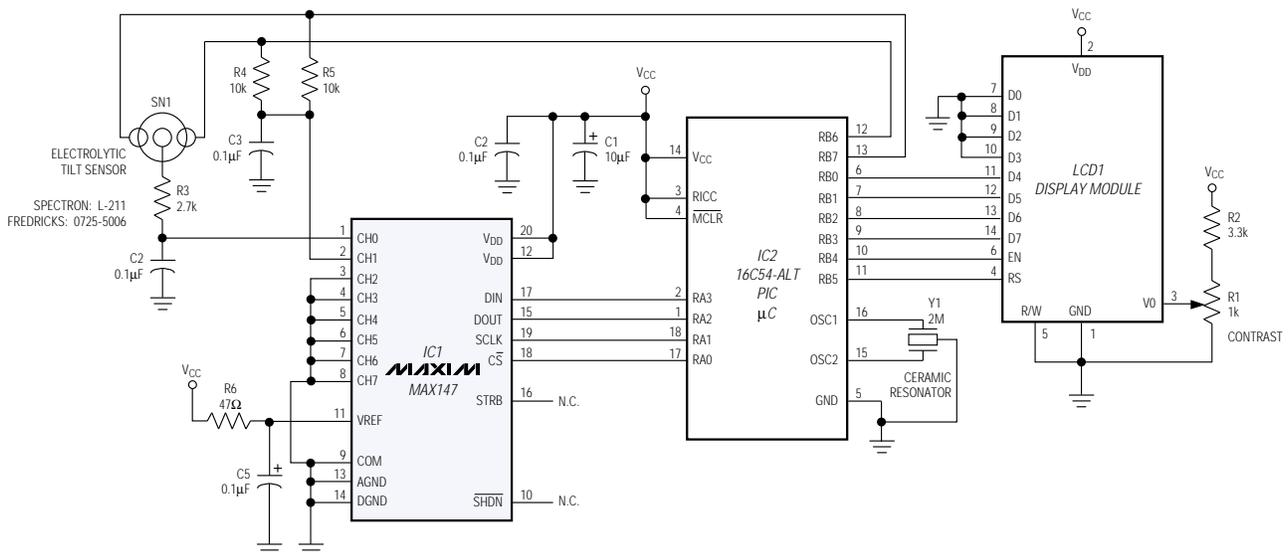


Figure 1. This tilt sensor is simple, accurate, inexpensive, and adjustment free.

The tilt signal on one channel and the reference (mid-level) signal on another are digitized by the ADC and fed to the μC . The AC drive dwells 10ms on each polarity, allowing about nine time constants for 12-bit settling before the A/D conversion. The converter's pseudo-differential input negates the absolute value of these signals ($\sim 1/2 V_{CC}$). Thus, the magnitude and polarity of channel 0 (with respect to channel 1) indicate the magnitude and direction of tilt. The tilt measurement is ratiometric and therefore relatively immune to large variations in the supply voltage (typically 0.2% of full scale per volt of supply change).

A measurement comprises two consecutive half-cycles: the μC first calculates the sensor-minus-reference value; it then applies an opposite-phase drive signal and calculates the reference-minus-sensor value. Subtracting these values produces twice the desired tilt value and negates the need for null adjustment by canceling any systematic offsets. The values are handled in software (see the software listing called "Adjustment-Free Inclinometer" under the Other Software category on Maxim's website at www.maxim-ic.com) as two's-complement quantities, and displayed on the liquid-crystal display (LCD) as integers. (The display in this system is included mainly for demonstrations.)

Miscellaneous observations

Though not implemented by the software provided, this system is capable of very-low-power operation. IC1 can be shut down between conversions; in this

state, it draws only $10\mu\text{A}$. While IC1 is shut down, pins 12 and 13 on the μC port should be written low to prevent DC current from damaging the sensor (consult the sensor's data sheet for the maximum DC current allowed). The μC 's internal watchdog can be set to wake up every second or so for a new measurement. Operating at a few measurements per second and replacing the LCD with Maxim's MAX7211 can lower the overall supply current to $100\mu\text{A}$.

The techniques previously described are compatible with most μC s and microprocessors (μP s), but the output structures of some μP s are unlike that of the Microchip PICTM. Most variants of the 8051, for example, have an open-drain output and pull-up resistor that exhibit unequal source and sink currents at the port pins. Ensure reliable operation for these variants by providing external CMOS inverters between the port pins and sensor. Design the power-up initialization and power-down conditions carefully to minimize DC current through the sensor.

Finally, these techniques can be expanded to accommodate dual-axis sensors by dedicating two more port pins for a second pair of force electrodes. The measurement procedure is nearly identical, except that the sensor pins for each axis must be alternately three-stated while making measurements on the other axis. This provision minimizes cross-axis interactions, which is a difficult task to accomplish with the more common analog techniques.

A related idea appeared in the 4/24/97 issue of EDN.

PIC is a trademark of Microchip Technology, Inc.

If the +5V supply fails or falls out of regulation (as defined by a 4.75V threshold determined by R3 and R4), the IC2 output goes high, disconnects the main supply by turning off Q3, turns on the backup supply by pulling IC1 out of shutdown, and connects the backup voltage and load by turning on Q1 and Q2. Q1 is chosen for low $R_{DS(ON)}$ (to minimize power dissipation), and Q2 is chosen for its ultra-low V_{GS} threshold (to ensure a reliable switchover to V_{BATT} when the main supply fails). The charge on C3 (present at all times, as mentioned previously) ensures a quick turn-on of the backup supply, and the charge on C4 supports the output voltage during switchovers between the +5V supply and V_{BATT} .

When the system is off (backup converter shut down and +5V supply absent), you can turn on the backup supply by momentarily pressing the "POWER ON"

pushbutton. (A connection from this switch to an I/O line gives on/off control to the processor and also allows the supply to send on/off signals to the processor.) Pressing the pushbutton turns on Q1, Q2, and IC1, enabling C4 to charge. When the pushbutton is released, R6's pull-up/latching effect takes over.

Q3's connection makes the drain more positive than the source. This unusual orientation allows the internal parasitic diode to conduct current when the +5V supply is connected, quickly charging C4 and providing power to IC2. (When IC2's output goes low, Q3 turns on, and its $R_{DS(ON)}$ shunts the parasitic diode.) Q3's low forward drop has a negligible effect on the main supply-voltage tolerance.

A related idea appeared in the 7/21/97 issue of Electronic Design.

DESIGN SHOWCASE

DAC-powered charge pump varies negative rail

The circuit shown in **Figure 1** provides a low-current, adjustable negative supply rail suitable for use as a sensor bias, liquid-crystal-display (LCD) contrast bias, or voltage-controlled-oscillator (VCO) tuning supply. By operating a charge-pump doubler from the output of a buffered digital-to-analog converter (DAC), the circuit avoids the customary approach involving clumsy level shifters based on op amps and discrete components.

IC1 is a dual, 8-bit DAC with serial input and buffered voltage outputs. Output impedances are 50Ω ; therefore, the DAC output in use drops about 50mV while providing the 1.1mA typically drawn by the IC2 charge pump. As the input code varies from 0 to 255, the DAC output ranges Rail-to-Rail®, changing approximately 40mV per step.

With a +5V input (V_{CC}) applied to IC1 and a -3V output from IC2, the code that produces the minimum allowable voltage to the charge pump (1.5V) is 80 (decimal). The charge pump draws 0.6mA and generates ± 2 times the voltage at its V_{CC}

terminal, producing $\pm 3V$ to $\pm 12V$ as its input ranges from 1.5V to 6V. (The positive and negative outputs can be used simultaneously.) The main power can go as low as 2.7V, producing a negative output slightly over -5V. The minimum code for this condition is about 140 (decimal).

To shut down the supply, simply write zeros to the DAC. The DAC itself has a shutdown mode that draws only $1\mu A$. To ensure a reliable start-up when bringing the system out of shutdown, write a value that powers the charge pump with a minimum of 2V. Note that a microcontroller (μC) with a pulse-width-modulation (PWM) output can eliminate the DAC altogether. For example, you can provide a variable V_{CC} to the charge pump by filtering a 20kHz PWM signal with a $270\Omega/3.3\mu F$ lowpass network. Be sure the μC 's port pin can supply the current with an acceptable voltage drop; if not, buffer it with a CMOS buffer or inverter such as the 74HC04.

A related idea appeared in the 7/21/97 issue of Electronic Design.

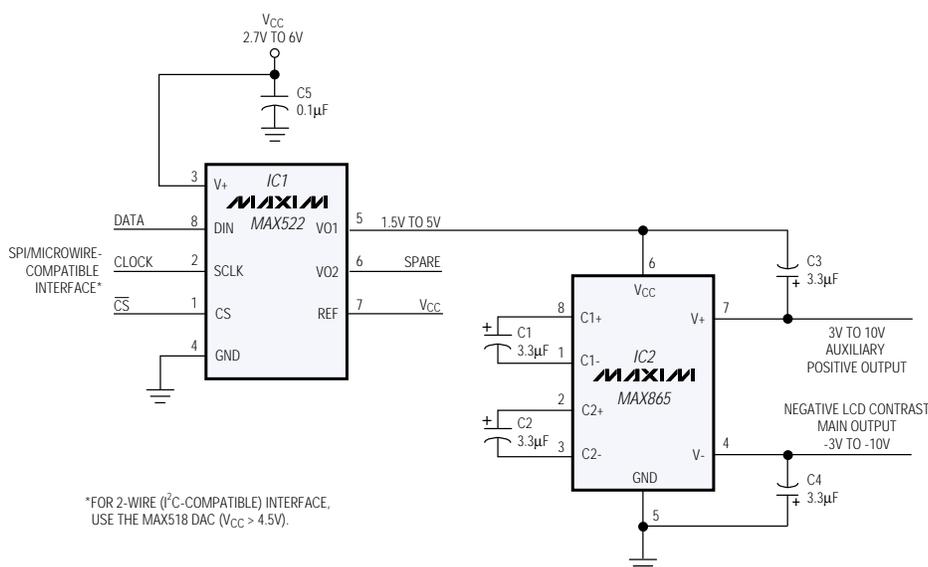


Figure 1. This adjustable negative supply consists of an inverting-doubler charge pump controlled by an 8-bit, serial-input DAC.

Rail-to-Rail is a registered trademark of Nippon Motorola Ltd.

DESIGN SHOWCASE

Boost controller drives buck converter

The usual way to step down from a low voltage to an even lower one is with a low-dropout (LDO) linear regulator. But in battery-powered systems, the LDO probably won't deliver the maximum energy available. A cell count chosen for near-dropout operation when the battery is empty applies too much voltage over most of the battery's discharge, and a cell count chosen for maximum efficiency over that range allows dropout well before the battery is empty.

One solution to this problem is the highly efficient buck DC-DC converter (**Figure 1**). This circuit can step down inputs as low as 2V to outputs as low as 1.25V, with efficiencies as high as 80% (**Figure 2**). Like an LDO, it works well at low input voltages. Unlike an LDO, its efficiency remains fairly high with inputs up to the allowable maximum (6.5V).

A step-up switching regulator (IC1) is made to step down with the addition of an external switching transistor (Q1). Via LX (pin 8), Q1 is driven by the IC's internal switching transistor: an open-drain, n-channel power MOSFET connected to ground. R2 limits the Q1 base current, and R1 turns Q1 off when

LX floats. The R1 and R2 values are chosen for maximum efficiency at light loads (1mA to 10mA), which limits the maximum available output current. Lower values for R1 and R2 allow higher output current, but cause the circuit to draw higher levels of quiescent current.

R3 and R4 determine output voltage, as shown in the following equation:

$$V_{OUT} = V_{REF} (R3 + R4) / R4$$

where $V_{REF} = 1.25V$.

The minimum output voltage is 1.25V (with R3 = 0 and R4 absent). R5 and R6 determine the threshold for low battery voltage in a similar manner. Input and output capacitors can be inexpensive electrolytic or tantalum types. For greatest efficiency, the inductor should be rated in excess of the desired output current, and it should have a reasonably low series resistance. Diode D1 should be a Schottky type, because losses are proportional to the diode's forward voltage, and this voltage is a substantial fraction of the output voltage.

A related idea appeared in the 6/5/97 issue of EDN.

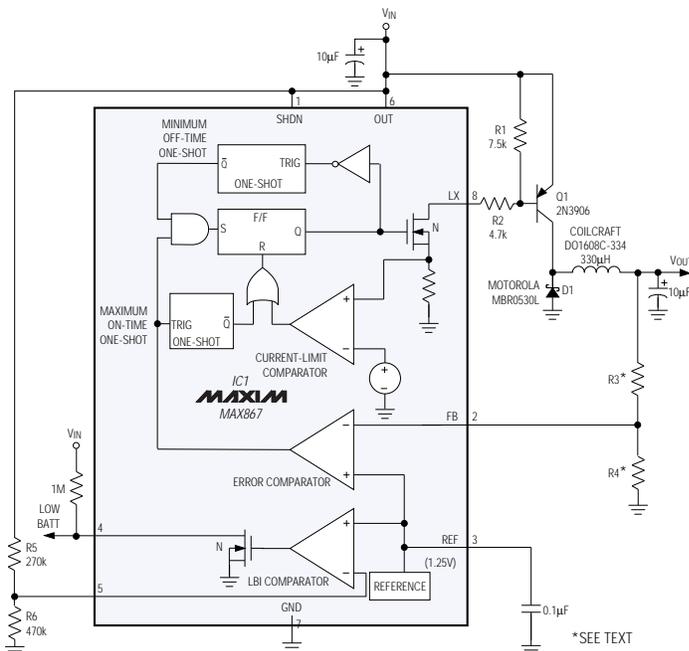


Figure 1. These external components enable a boost-controller IC to implement a low-voltage buck-regulator circuit.

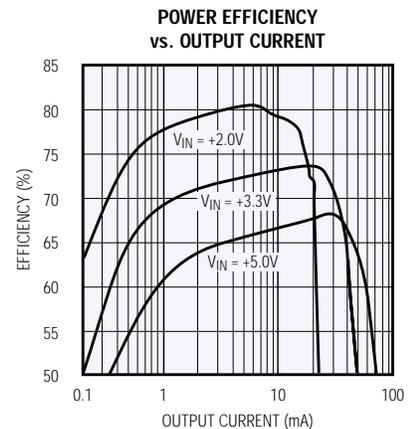


Figure 2. The conversion efficiency of the circuit in Figure 1 varies with output current as shown.

NEW PRODUCTS

IF undersampling CODEC combines digitizing ADC with reconstruction DAC

The MAX1005 intermediate-frequency (IF) undersampling CODEC provides an interface between the analog and digital portions of a PWT1900* communications system. This device includes a 5-bit analog-to-digital converter (ADC) for receiver-IF bandpass sampling, a 7-bit digital-to-analog converter (DAC) for reconstructing an analog IF subcarrier, and a separate, low-noise bandgap reference for each.

The ADC's 15Msps conversion rate enables 10x oversampling of a 1.5MHz signal. However, its wide input bandwidth (15MHz) allows IF undersampling in excess of 10.7MHz. The DAC has very low glitch energy and high spurious-free

dynamic range to minimize the transmission of unwanted spurious signals.

The MAX1005 can operate from either a single power supply or from separate analog and digital supplies, and with independent voltages ranging from +2.7V to +5.5V. These might include, for example, an unregulated analog supply of +5.5V and a regulated digital supply as low as +2.7V. This flexibility allows operation directly from a battery, even when the battery is being charged, thereby eliminating the noise associated with switching regulators and saving the power otherwise lost in linear regulators.

Operating modes include transmit (DAC active), receive (ADC active), and full shutdown, in which the supply current drops below 1 μ A. Because the wake-up time from partial shutdown is only 2.5 μ s, the MAX1005 can save power during short intervals of idle time.

The MAX1005 is available in a tiny 16-pin QSOP package specified for the commercial (0°C to +70°C) or extended-industrial (-40°C to +85°C) temperature range. Prices start at \$2.96 (1000 up, FOB USA).

*PWT1900 is a PCS air-interface standard for the U.S. Based on the proven DECT technology, the PWT1900 standard is suitable for use in toll-quality wireless PBX, PCS, and WLL applications.

+2.7V, 12-bit/10-bit ADCs have internal reference

The MAX1240/MAX1242 12-bit/10-bit ADCs feature low-power operation and an internal reference. Pin and software compatible, they each combine a track/hold, ADC, reference, clock, and serial interface in an 8-pin SO package.

These converters operate from a +2.7V to +3.6V single supply. They draw less than 2mA (including reference current) at a 73ksps (max) sampling rate. The supply currents drop to only 2 μ A in shutdown. The 3-wire serial interface is compatible with SPI™/QSPI™ and Microwire™ synchronous-serial standards.

The MAX1240 (offered in three grades) and MAX1242 (offered in two grades) are available in 8-pin SO and DIP packages. Prices start at \$2.75 for the MAX1242 and \$3.85 for the MAX1240 (1000 up, FOB USA).

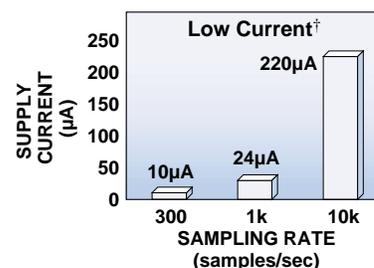
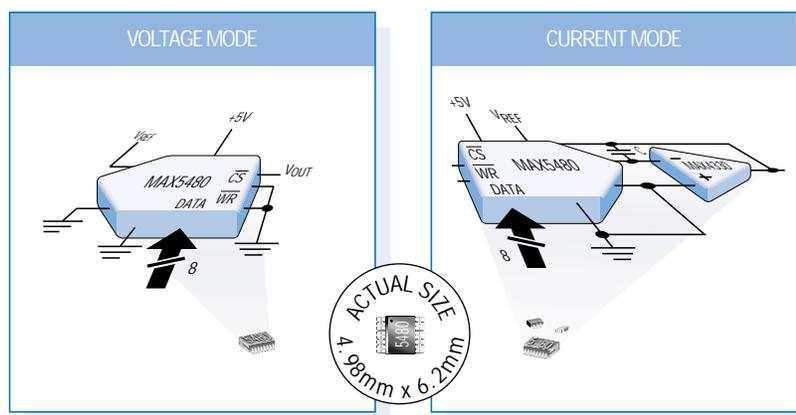
SPI and QSPI are trademarks of Motorola, Inc. Microwire is a trademark of National Semiconductor Corp.

8-bit latched parallel DAC resides in 16-pin QSOP package

The MAX5480 is an 8-bit, parallel-input, CMOS DAC that interfaces directly with most microprocessors. Its internal input latches make the DAC interface similar to a random-access-memory write cycle, in which the only control inputs are CS and WR.

Specifications guaranteed over temperature include monotonicity, $\pm 1/2$ LSB linearity, and 2.5mW power consumption. The MAX5480 operates from a single +5V supply and draws maximum supply currents of 100 μ A at +25°C and 500 μ A over temperature. It offers both current-output and voltage-output operation.

The MAX5480 is available in 16-pin CERDIP and QSOP packages, and in versions specified for three different temperature grades. Prices start at \$1.35 (1000 up, FOB USA).



† V_{REF} = V_{DD}. Using the internal reference, the supply current at 1ksps is 139 μ A.

NEW PRODUCTS

Low-cost, high-speed, single-supply SOT23 buffers have rail-to-rail outputs

Members of the MAX4014 family of precision, closed-loop, high-speed buffers provide a high slew rate (600V/μs), wide bandwidth (200MHz at -3dB), high output current (±120mA), and low gain/phase error (0.02%/0.02°), while drawing only 5.5mA of quiescent current per amplifier. The outputs swing rail-to-rail, and the input common-mode voltage ranges extend 200mV beyond the negative supply rail.

MAX4014 family buffers are well suited for use in video, communications, instrumentation, and other low-power/low-voltage applications requiring wide bandwidth. Operating from a +3.3V to +10V single supply or a ±1.65V to ±5V dual supply, they exhibit only 10nV/√Hz and 1.3pA/√Hz of input noise at the inverting or noninverting input. The triple-buffer MAX4019 has a disable feature that reduces supply current to 350μA.

The MAX4014 (single), MAX4017 (dual), MAX4019 (triple), and MAX4022 (quad) are available in space-saving SOT23-5, μMAX, or QSOP packages. Prices start at \$0.98 (1000 up, FOB USA).

High-accuracy, current-limited switch eases power-supply requirement by 50%

The MAX869L current-limited power switch features low on-resistance (only 35mΩ at 5V) and a current limit that is ±20% accurate and adjustable from 400mA to 2.5A. It protects systems from short-circuit and overload faults. In Universal Serial Bus (USB) applications, for example, such faults at a card slot or plug-in port can pull the main supply voltage below its minimum operating level.

Tight tolerance on the output current limit is critical to keeping the main power supply simple and inexpensive. To ensure a minimum continuous current of 2A, for example, the MAX869L maintains a nominal 2.5A with a maximum of 3A. Similar parts from other suppliers can guarantee only ±50% accuracy; therefore, they must maintain a nominal 4A and a maximum 6A. Thus, better accuracy in the MAX869L reduces the power-supply requirement by 50% (from 6A to 3A).

The MAX869L includes thermal-overload protection, and its current-limit loop features a fast, 4μs response that prevents system glitches and resets during hot plug-ins, when heavy capacitive loads can cause a momentary short circuit. When the MAX869L goes into current limit or thermal overload, its logic FAULT output alerts a microprocessor.

A 2.7V to 5.5V input range makes the MAX869L ideal for both 3V and 5V systems. It features a very low quiescent current (12μA) that drops to only 0.01μA in the OFF state. Typical applications include notebook and hand-held computers with slots and ports for the USB, PCMCIA, and CardBus, as well as power ports for peripheral devices.

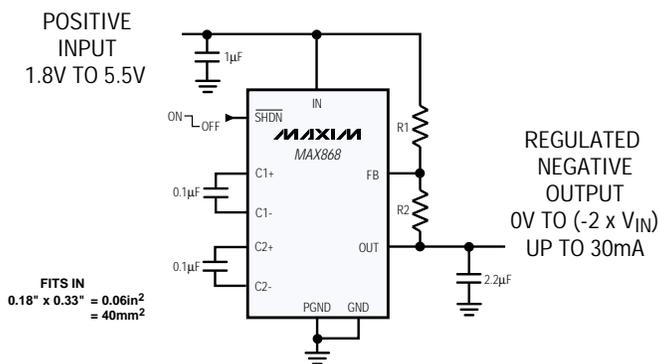
The MAX869L is the newest member in a family of high-side, p-channel, MOSFET power switches (MAX890L–MAX895L). The MAX869L is available in a 16-pin QSOP package (same board area as an 8-pin SO). Prices for the MAX869L start at \$2.13 (1000 up, FOB USA).

Regulated supply (3VIN, -5VOUT) is 0.06in² by 1.11mm high

The MAX868 is an adjustable, regulated, switched-capacitor voltage converter that inverts, then doubles inputs of 1.8V to 5.5V. As a compact, low-cost means for generating negative supply voltages equal to -2VIN, this 30mA charge-pump device reduces cost, board area, and height by replacing inductor-based DC-DC converters. Typical applications include cell phones, small LCD panels, and PCMCIA cards.

Regulation is achieved by gating the 450kHz charge-pump oscillator to keep the output voltage constant. This on-demand switching scheme provides excellent light-load efficiency and generates output currents as high as 30mA under full load. For operation, the MAX868 requires four ceramic capacitors and, to set the output voltage, two external resistors. Optimized for battery-operated equipment, the MAX868 features a quiescent supply current of only 30μA and a logic-controlled shutdown pin that turns off the charge pump and reduces the total current to less than 1μA.

The MAX868 is available in a 10-pin μMAX package (only 1.11mm high, covering half the area of an 8-pin SO). Prices start at \$1.75 (1000 up, FOB USA).



NEW PRODUCTS

Precision PWM buck controller optimized for next-generation notebook CPUs

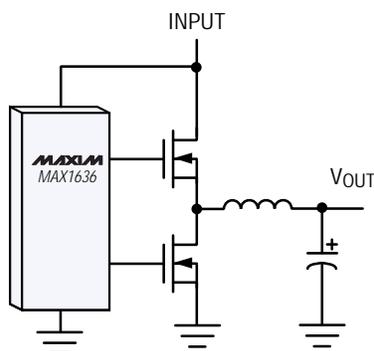
The low-voltage MAX1636 pulse-width modulation (PWM) controller generates precisely regulated CPU supply voltages from the high-voltage battery in a notebook computer. As a fixed-frequency, current-mode PWM controller, the MAX1636 provides fast transient response, low supply current, tight load regulation, and the tight output accuracy required by today's CPUs. Other applications include battery chargers, inverters, and boost-topology circuits.

The MAX1636's combination of a low-drift reference, slow integrator loop, and fast current-mode loop provides an exceptional $\pm 1\%$ output accuracy, including all conditions of line and load. To protect the output from overvoltage, a crowbar circuit turns on the low-side MOSFET (in less than $1\mu\text{s}$, with 100% duty cycle) when the feedback signal goes high by more than 7%. As additional fault protection, a catastrophic undervoltage detector shuts down the PWM if the output

fails to come into regulation within a preset time. The quiescent power dissipation is only 2mW (max).

The MAX1636 accepts 4.5V to 30V inputs and generates a 1.1V to 5.5V adjustable output voltage. It includes a 5V/25mA linear regulator (off in shutdown but on in standby mode) that provides a gate-drive supply for the low-side external MOSFET. For a similar device that does not include a linear regulator and comes in a smaller, 16-pin package, refer to the MAX1637.

The MAX1636 is available in a 20-pin SSOP specified for the extended-industrial temperature range (-40°C to $+85^{\circ}\text{C}$). Prices start at \$3.95 (1000 up, FOB USA).



Remote/local temperature sensor has SMBus serial interface

The MAX1617 is a precise digital thermometer that reports the temperature of its own package in addition to that of a remote sensor. The remote sensor—an easily mounted, diode-connected npn transistor such as the low-cost 2N3904—can replace a conventional thermistor or thermocouple. With Maxim's patented measurement circuitry, such transistors from multiple manufacturers can provide $\pm 3\%$ accuracy without calibration. The remote channel can also measure the temperature of any IC (such as a microprocessor) that includes an accessible diode-connected transistor.

To read temperature data and program the alarm thresholds, the MAX1617 accepts standard write-byte, read-byte, and receive-byte commands via a 2-wire serial interface called the System Management Bus (SMBus™). The data format is seven bits plus sign, twos-complement, in which each LSB represents 1°C . Conversion rate (and therefore current drain) is programmed by the user, who also programs the under- and over-temperature alarms and sets the device for single-shot or continuous measurements.

The MAX1617 operates from a +3V to +5.5V supply and draws only $3\mu\text{A}$ (typ) in standby mode. It comes in a 16-pin QSOP package specified for the military temperature range (-55°C to $+125^{\circ}\text{C}$). Contact the factory for pricing.

SMBus is a trademark of Intel Corp.

Switch-mode controllers provide digitally adjustable LCD-bias voltage

The MAX1620/MAX1621 digitally adjustable LCD-bias supplies come in ultra-small QSOP packages and operate with small, low-profile external components. Each is suitable for use in notebook and palmtop computers, personal digital assistants, and portable data-collection terminals. Operating from +1.8V to +20V battery voltages, they produce positive or negative output voltages of $\pm 27\text{V}$.

External resistors set the desired maximum and minimum output voltages, and a high or low connection at the POL terminal sets the output polarity. To adjust over this range, employ either an external potentiometer or digital software control via the internal 5-bit digital-to-analog converter. The MAX1620 allows up/down digital signaling for this purpose; the MAX1621 provides an interface for control via the 2-wire-serial SMBus.

In typical applications, the MAX1620/MAX1621 are powered from the display's +3V to +5.5V logic supply and draw $150\mu\text{A}$ ($250\mu\text{A}$ max). Connecting the SHDN input to this supply protects the display: a loss of supply voltage triggers shutdown, which removes bias voltage from the display and drops the chip's quiescent current to $10\mu\text{A}$ (max). The external power switch can be an n-channel MOSFET or a low-cost npn transistor.

The MAX1620/MAX1621 are available in 16-pin QSOP packages specified for the extended-industrial temperature range (-40°C to $+85^{\circ}\text{C}$). Prices start at \$1.99 (1000 up, FOB USA).

NEW PRODUCTS

Low-dropout linear regulators generate 3.3V, 5V, or adjustable outputs

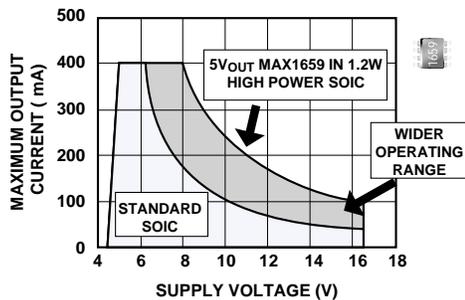
The MAX1658/MAX1659 linear regulators have ultra-low supply currents and low dropout voltages that maximize battery life. Their Dual Mode™ operation provides either preset outputs of 3.3V (MAX1658) or 5V (MAX1659), or adjustable outputs from 1.25V to 16V. The input voltage range is from 2.7V to 16.5V.

Output current capability is 350mA, with typical dropout voltages of 650mV (MAX1658) and 490mV (MAX1659).

Internal p-channel MOSFET pass transistors enable each device to maintain a low quiescent supply current (30μA) while providing output currents from zero to 350mA, even in dropout. Maximum supply current in shutdown is 1μA. Other MAX1658/MAX1659 features include reverse-battery protection, short-circuit protection, and thermal shutdown.

The MAX1658/MAX1659 are available in a special 8-pin SO package, specified for the extended-industrial temperature range (-40°C to +85°C), with a high power rating (1.8W) that supports compact applications. Prices start at \$1.95 (1000 up, FOB USA).

Dual Mode is a trademark of Maxim Integrated Products.



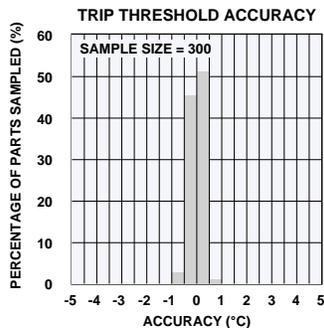
Low-cost, SOT temperature switches feature logic outputs

All of the MAX6501–MAX6504 low-cost temperature switches include a comparator with two temperature-dependent voltage references. They draw 30μA (typical) from a single supply voltage of +2.7V to +5.5V. The internal temperature-trip thresholds (designated by the part number suffix) are factory set in 10°C increments from -45°C to +115°C. Accuracy is ±0.5°C typical (±4°C max) over the specified temperature range, and hysteresis is pin-selectable as +2°C or +10°C. No external components are required.

The MAX6501/MAX6503 have active-low, open-drain outputs suitable for driving a microprocessor's reset input, and the

MAX6502/MAX6504 have active-high, push/pull outputs suitable for driving fan-control logic. All devices assert a logic output when the measured temperature crosses the factory-set threshold for hot temperature (MAX6501/MAX6503) or cold temperature (MAX6502/MAX6504).

MAX6501–MAX6504 monitors are available in 5-pin SOT23 packages specified for the military temperature range (-55°C to +125°C). Prices start at \$0.50 (10,000 up, FOB USA).



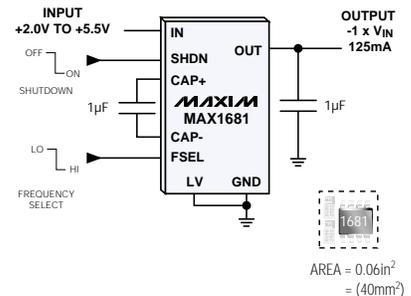
1MHz charge pumps generate 125mA

The MAX1680/MAX1681 are high-frequency, switched-capacitor voltage converters that supply up to 125mA of output current when doubling or inverting 2.0V to 5.5V inputs. They offer the most compact method available for doubling an input voltage or generating a negative supply for amplifier and analog-measurement circuits. By replacing inductor-based DC-DC converters, the MAX1680/MAX1681 reduce cost, board area, and height.

The MAX1681 allows users to select a 500kHz or 1MHz operating frequency, and the MAX1680 allows a choice between 125kHz and 250kHz. These four frequencies enable the designer to adjust quiescent supply current versus external capacitor size at a given output current.

At 1MHz, the MAX1681 exhibits only 3.5Ω of output resistance when operating with 1μF external capacitors. The MAX1680 requires 10μF capacitors to maintain this output resistance. Using 1206-size 1μF capacitors, the MAX1681 circuit's board area is only 0.06in² (40mm²). Both devices feature a logic-controlled shutdown that turns off the charge pump and lowers the quiescent current to less than 1μA.

The MAX1680/MAX1681 are available in an 8-pin SO package with prices starting at \$2.05 (1000 up, FOB USA).



NEW PRODUCTS

250kbps RS-232 transceiver adds power-on reset function

The MAX3320 is a dual RS-232 transceiver that provides supply-voltage monitoring and automatic power-down (after a 30-second interval with no valid data transitions). Applications include notebook and palmtop computers, high-speed modems, and printers.

The MAX3320's power-on reset function asserts an active-low reset when V_{CC} declines below a preset threshold. It maintains the reset for at least 140ms after V_{CC} returns above the threshold. The reset comparator ignores fast V_{CC} transients, and the reset signals are guaranteed correct for V_{CC} levels down to 1V. MAX3320 threshold variants (indicated by suffix letter) accommodate a variety of supply voltages.

The dual transceiver (two transmitters and two receivers) includes a dual charge pump and a proprietary transmitter output stage whose low dropout enables valid RS-232 levels for supply voltages between +3V and +5.5V. It requires only four small 0.1 μ F capacitors for operation and guarantees data rates as high as 250kbps.

A shutdown mode reduces power consumption and extends battery life in portable systems by lowering the MAX3320 supply current to 4 μ A. The receivers and power-on-reset function remain active in shutdown to monitor modems and other external devices. Maxim's AutoShutdown Plus™ technology, when enabled, places the MAX3320 in shutdown when 30 seconds elapse without a valid signal transition on the receiver-input lines.

The MAX3320 is available in a 20-pin SSOP package, in versions specified for the commercial (0°C to +70°C) or extended-industrial temperature range (-40°C to +85°C). Prices start at \$1.85 (1000 up, FOB USA).

AutoShutdown and AutoShutdown Plus are trademarks of Maxim Integrated Products.

1Mbps RS-232 transceiver has AutoShutdown and ± 15 kV ESD protection

The new MAX3243E/MAX3244E/MAX3245E are 1Mbps, RS-232 communications transceivers. Each is a complete serial port consisting of three drivers and five receivers. These devices are intended for use in notebook or subnotebook computers, and are guaranteed to drive a mouse. Features include automatic shutdown and wake-up, high maximum data rate, and enhanced protection from electrostatic discharge (ESD). All transmitter outputs and receiver inputs are ESD protected to ± 15 kV using the Human Body Model or the IEC 1000-4-2 Air-Gap Discharge method, and to ± 8 kV using the IEC 1000-4-2 Contact Discharge method.

The transceivers' regulated dual charge pump and proprietary low-dropout transmitter outputs enable true RS-232 performance while operating from a +3.0V to +5.5V single supply. Supply current is only 300 μ A. RS-232 output levels are maintained by the MAX3243E/MAX3244E at data rates to 250kbps, and by the MAX3245E (which includes Maxim's Megabaud™ feature) to data rates as high as 1Mbps.

MAX3243E/MAX3244E/MAX3245E logic-controlled shutdown lowers the supply current to 1 μ A. AutoShutdown™ and AutoShutdown Plus enable these systems to save power automatically, without changing the BIOS or the operating system. AutoShutdown, for example, shuts down the MAX3243E if the RS-232 cable is disconnected or if any connected peripheral is turned off. This device turns back on when a valid level appears at any receiver input. AutoShutdown Plus devices (MAX3244E/MAX3245E) enter shutdown 30 seconds after a cable is disconnected or a peripheral becomes idle, and they resume operation when a valid edge appears at any transmitter or receiver input.

MAX3243E/MAX3244E/MAX3245E transceivers are available in 28-pin SO and SSOP packages. Prices start at \$3.82 (1000 up, FOB USA).

MegaBaud is a trademark of Maxim Integrated Products.

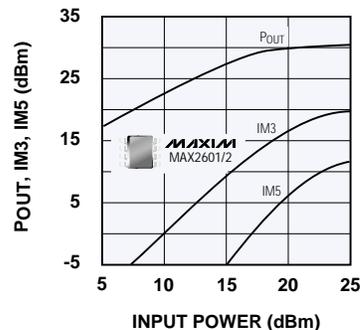
+3V, 1W RF power transistors ideal for 900MHz applications

The MAX2601/MAX2602 are low-voltage bipolar power transistors. Their high gain and efficiency make them ideal for the final stage of a class-C or class-AB RF amplifier, whether in discrete or module form. They reduce cost and save space by eliminating the need for drain switches and negative-bias generators.

The MAX2601/MAX2602 exhibit 11.5dB gain while producing 1W of RF power at 900MHz from a 3.6V supply voltage. The collector efficiency is 58%, and the second- and third-harmonic suppression is 43dBc. These devices withstand load mismatch conditions ($V_{SWR} = 8:1$ at all angles, with $V_{CC} = 5.5$ V) without exhibiting spurious oscillations or excessive power draw. In addition to the power transistor, the MAX2602 also includes a thermal- and process-matched diode, which allows simple and accurate biasing with just one external resistor.

The MAX2601/MAX2602 are available in a thermally enhanced, 8-pin SO package. Prices start at \$2.32 for the MAX2601 and \$2.38 for the MAX2602 (1000 up, FOB USA).

TWO-TONE OUTPUT POWER vs. INPUT POWER



NEW PRODUCTS

Direct-conversion IC tunes DBS television signals

The MAX2102 direct-conversion tuner is intended for use in set-top boxes for direct-broadcast satellite (DBS) digital television. In comparison with intermediate-frequency (IF) architectures, this device reduces cost by eliminating the IF mixer, IF local oscillator, and SAW filter.

The MAX2102 downconverts 950MHz to 2150MHz L-band signals to the desired baseband. It includes a low-noise amplifier, automatic gain control (AGC) amplifier, two downconverter mixers, an oscillator buffer with dual-modulus prescaler that divides by 64 or 65, a 90° quadrature (Q) generator, and separate baseband amplifiers for the in-phase (I) and Q outputs.

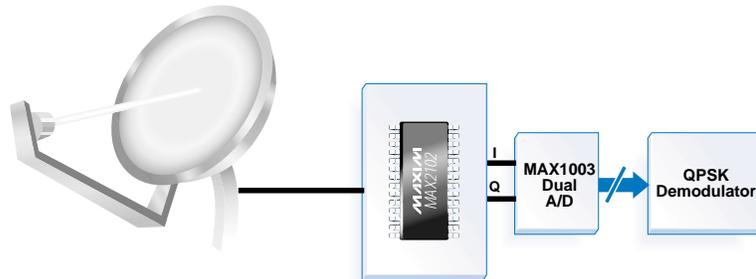
The MAX2102 was designed with Maxim's high-frequency bipolar process (GST-2, with $f_T = 27\text{GHz}$). It achieves $\pm 3^\circ$ Q phase accuracy and $>0.5\%$ gain mismatch between I and Q channels over

the entire frequency range. Operating from a single +5V supply, it provides an AGC control range of more than 50dB for -19dBm to -69dBm input signals. This large AGC range accommodates rainfall attenuation effects, different cable lengths, and less-than-perfect alignment of the DBS parabolic dish antenna.

At 1450MHz, the noise figure is 13.2dB, and the input third-order intercept point (6.5dBm) enables a single discrete preamplifier to serve as the interface to a 75 Ω cable. Internal offset-correction

amplifiers (when not disabled) automatically remove any DC offset present in the baseband amplifiers. Channel selection in the baseband is performed by discrete, low-cost LC filters, typically with a 5th- or 7th-order lowpass response.

To support MAX2102 customers, Maxim offers a preassembled evaluation kit (MAX2102EVKIT) and tuner-design assistance. The MAX2102 is available from \$5.00 (1000 up, FOB USA) in a 28-pin SO package.



Low-voltage IF transceiver includes limiter and RSSI

The MAX2511 is an IF transceiver that incorporates a multitude of radio functions in an ultra-small package. Applications include PCS systems such as PWT1900, PACS, PHS, and DECT phones and base stations. In these systems, the MAX2511 performs all transmit and receive functions from first IF to a 10.7MHz second IF. The MAX2511 also serves as a highly integrated front-end radio transceiver for use in applications with 200MHz to 440MHz carrier frequencies, such as ISM transceivers.

The MAX2511 receiver section features an image-reject downconverter with 34dB of image suppression, followed by a wide-dynamic-range IF buffer that drives an off-chip IF filter. Next is a limiting amplifier with differential outputs that boosts the signal to 1Vp-p. The

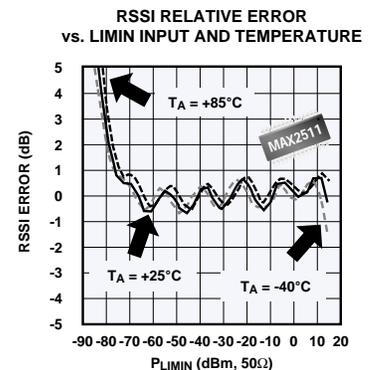
received-signal-strength indicator derived from the limiting amplifier (RSSI output) has more than 90dB dynamic range and excellent linearity: its guaranteed maximum relative error is $\pm 2\text{dB}$.

The transmitter section includes image-reject upconversion and a variable-gain, 0dBm output-buffer amplifier. To extend battery life, the amplifier's unique biasing scheme adjusts current draw to the minimum necessary to sustain the desired output power level.

An external tank circuit completes the internal voltage-controlled oscillator (VCO), which includes a buffer for driving an external phase-locked loop. Power to the VCO is internally regulated to ensure a constant-frequency output. The MAX2511's +2.7V to +5.5V operating supply voltage enables direct connection to a 3-cell battery, and four power-control settings enable advanced system power management. A shutdown mode lowers the chip's supply current to below 2 μA .

As an alternative for applications that do not require image rejection at the transmitter or receiver, Maxim will soon introduce the MAX2510*. It includes most of the other features found in the MAX2511 (limiter, RSSI, control functions, etc.). The MAX2511 comes in a 28-pin QSOP package specified for the extended-industrial temperature range (-40°C to +85°C). Prices for the MAX2511 start at \$5.94 (1000 up, FOB USA).

*Future product—contact factory for availability.



NEW PRODUCTS

Image-reject RF transceiver ideal for low-cost 900MHz radios

The MAX2420 is a highly integrated, front-end IC designed to reduce cost in 900MHz cordless telephones, wireless modems, and RF transceivers. Its direct-conversion image-reject mixers, which eliminate external filters and extra frequency conversion, save at least \$2.00 per unit in each of these applications.

Unlike conventional RF transceivers, the MAX2420 includes active image-reject mixers. On the receiver side, image rejection allows a very low IF without the need for complex filters, which also eliminates the need for a second frequency conversion. On the transmit side of a frequency-hopping system, the image-reject mixer enables direct upconversion of a digitally synthesized transmit signal. This capability reduces cost and saves space by eliminating trims and additional frequency-conversion stages.

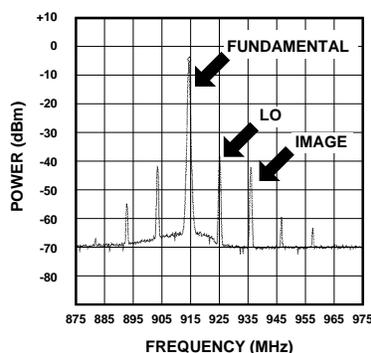
The MAX2420's mixers have a 35dB (typ) image rejection. The low-noise amplifier with 1.8dB noise figure allows a combined downconverter noise figure of

just 4dB with an input third-order intercept point (IP3) of -17dBm. Low-noise amplifier gain is adjustable to increase the receiver dynamic range (up to 2dBm input IP3).

The MAX2420 is optimized for 10.7MHz transmit and receive IFs; future versions will be optimized for 45MHz, 70MHz, and 110MHz IFs. Future members of this family will replace the transmit image-reject mixer with a balanced mixer, which can be used as a balanced modulator or local-oscillator buffer. Receive-only versions will also be available.

The MAX2420 operates from a +2.7V to +4.8V single supply. It is available in a 28-pin SSOP package, with prices starting at \$4.49 (1000 up, FOB USA).

TRANSMITTER OUTPUT SPECTRUM



Lowest phase-noise RF oscillator replaces VCO modules

The MAX2620 is a low-noise oscillator that operates from 10MHz to 1050MHz. When the MAX2620 is properly mated with an external varactor-tuned tank circuit, its typical phase noise is only -110dBc/Hz at 25kHz offset from a 900MHz carrier.

This low-noise capability, combined with +2.7V to +5.25V single-supply operation, makes the MAX2620 an ideal choice for next-generation analog and digital cellular phones, 900MHz cordless phones, land-mobile radio, and narrowband PCS systems that operate from three nickel-cadmium/nickel-metal hydride cells or a single lithium-ion cell.

The MAX2620 includes a low-noise transistor, two buffer amplifiers, biasing circuitry, and a power-save capability that lowers the supply current from 9mA during operation to 0.1 μ A in shutdown mode. Oscillation frequency is set by an external varactor-controlled ceramic resonator or LC tank. The MAX2620 boasts minimal frequency pushing—only 71kHz per volt of supply change—making it less sensitive to the sudden supply-voltage changes common in TDMA systems.

The MAX2620 comes in an 8-pin μ MAX package specified for the extended-industrial temperature range (-40°C to +85°C). Prices start at \$1.98 (1000 up, FOB USA).

+3.3V clock-recovery/data-retiming IC suits 622Mbps SDH/SONET receivers

The MAX3675 is a clock-recovery and data-retiming IC for SDH/SONET and ATM applications. Designed for use in 622Mbps, NRZ-serial-data receivers, it has selectable dual inputs that accept either small-signal analog or differential-PECL data. A high-gain limiting amplifier at the analog input accepts 3.6mVp-p to 1.2Vp-p signals.

Operating from a single +3.3V supply, the MAX3675 consumes only 215mW. It complies with ANSI, ITU, and Bellcore specifications for Type-A regenerators. No external reference clock is required. A fully integrated phase-locked loop with loss-of-lock monitor tracks the external signal. Clock and data outputs are in differential-PECL format. The limiting amplifier generates a loss-of-power signal, for which an internal bandgap reference lets you set the trip point independently of the supply voltage.

The limiting amplifier also includes an extremely fast, logarithmic-signal power detector that provides a received-signal-strength indicator (RSSI). The power detector acts as a broadband power meter, detecting the total RMS power of all signals in the passband. Temperature and power-supply independent, its RSSI voltage varies linearly in decibels from 1.35V to 2.4V, for -50dBm to -10dBm (2mVp-p to 200mVp-p) input power levels. Also included is a fully integrated input-offset correction loop that requires no external filter components.

The MAX3675 is available in die form and in a 32-pin TQFP package, both specified for the extended-industrial temperature range (-40°C to +85°C). Pricing for the packaged part (1000 up, FOB USA) starts at \$44.39.